REPORT MDC E2209

NASA CR-

(NASA-CR-160573) ULTRAMICROWAVE COMMUNICATIONS SYSTEM, PHASE 2 Final Report (McDonnell-Douglas Astronautics Co.) 33 p HC A03/MF A01 CSCL 17B Unclas G3/32 18026

N80-22545

# ULTRAMICROWAVE COMMUNICATIONS SYSTEM PHASE II

## FINAL REPORT

MCDONNELL DOUGLAS ASTRONAUTICS COMPANY - ST. LOUIS DIVISION



MCDONNELL DOUGLAS CORPORATION

COPY NO. 10

# ULTRAMICROWAVE COMMUNICATIONS SYSTEM PHASE II

5 JANUARY 1980

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## FINAL REPORT

SUBMITTED TO:
NATIONAL AERONAUTICS AND SPACE ADMINISTRATION
LYNDON B. JOHNSON CENTER
HOUSTON, TEXAS 77058

IN RESPONSE TO CONTRACT NO. NAS-9-15617

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## 1. INTRODUCTION AND SUMMARY

Phase II of the Ultramicrowave Communications System program had five main tasks: refining and analyzing the preliminary design of the RF subsystem, performing a packaging study of a complete communications system, developing a mockup design of the Phase III test system, procuring the required active components for the mockup demonstration, and finalizing the assembly and test plan requirements for Phase III.

Communications system design was completed and reviewed. Minor changes were made in the Phase I design of the RF subsystem in order to make it more cost effective and to increase design flexibility with regards to demonstration of data rate capability and maximization of efficiency. System design activities have also identified the techniques and procedures to generate and monitor high data rate test signals. Although differential bi-phase demodulation is the proposed method for this system, other demodulating schemes for the RF subsystem have been studied with a positive conclusion.

The mockup and packaging designs were performed by the Electronics Equipment Development department of MDAC-STL. Component layout and interconnection constraints were determined as well as design drawings for dummy parts of the system. Compactness for the transmitter/receiver configuration has been demonstrated and documented.

Positive results were achieved from a study to determine the possibility of adding a low-cost radar option to the transceiver system. The communications program has the advantage that new technology signal processing devices, (microprocessors, SAW components, etc.) can be readily interfaced with the existing RF subsystem to produce a short range radar. The four principal reasons for using ultramicrowave frequencies; private spectrum allocations, size compactness, high

data rate, and jamming resistance clearly enhance the attractiveness of the development of millimeter wave radar systems.

# 2. DESIGN DEVELOPMENT

A thorough design study for the system to be used in the Phase II demonstration plus studies for future systems were completed. A summary of the design evaluation is given below.

## 2.1 System Design

This section summarizes the work performed on the RF and signal processing subsystem as well as conclusions about alternate modulation and demodulation methods.

### 2.1.1 Summary of RF Subsystem Design and Performance Analysis

RF subsystem design activities centered around the refining of the recommended preliminary design, illustrated in figure 1. A restudy of the proposed antenna design has lead to a substitution of a TRG 802-6 prime focus antenna for the component parts described in the technical proposal of Phase II (MDC E2060). It is more cost effective to use complete manufactured antennas rather than assemble them ourselves from component parts. Also recommendations indicate that the surface tolerance be held to less than $\lambda/50$ for reflectors[1]. At 105 GHz $\lambda = .1125$ inches so $\lambda/50 = 2.25 \times 10^{-3}$ inches. The tolerance of the previously proposed antenna is .01 inches, clearly too high for operation in this frequency range. However the tolerance of the 802-6 antenna is .001 inches, which is sufficient for operation at 105 GHz.

Further analysis has lead to a substitution of a Hughes 47136H-1105 micrometer tuned CW IMPATT source for the originally proposed Hughes 47176H-1005 IMPATT source. The micrometer tuned source provides more flexibility in the choice of a carrier frequency which will aid in maximizing the efficiency (and data rate capability) of the system. Also it will be useful for systems design at other

than 105 GHz. Table 1 contains the performance analysis of the RF demonstration system for Phase III as well as other possible subsystems.

FIGURE 1: PRELIMINARY DESIGN FOR RF SUBSYSTEM

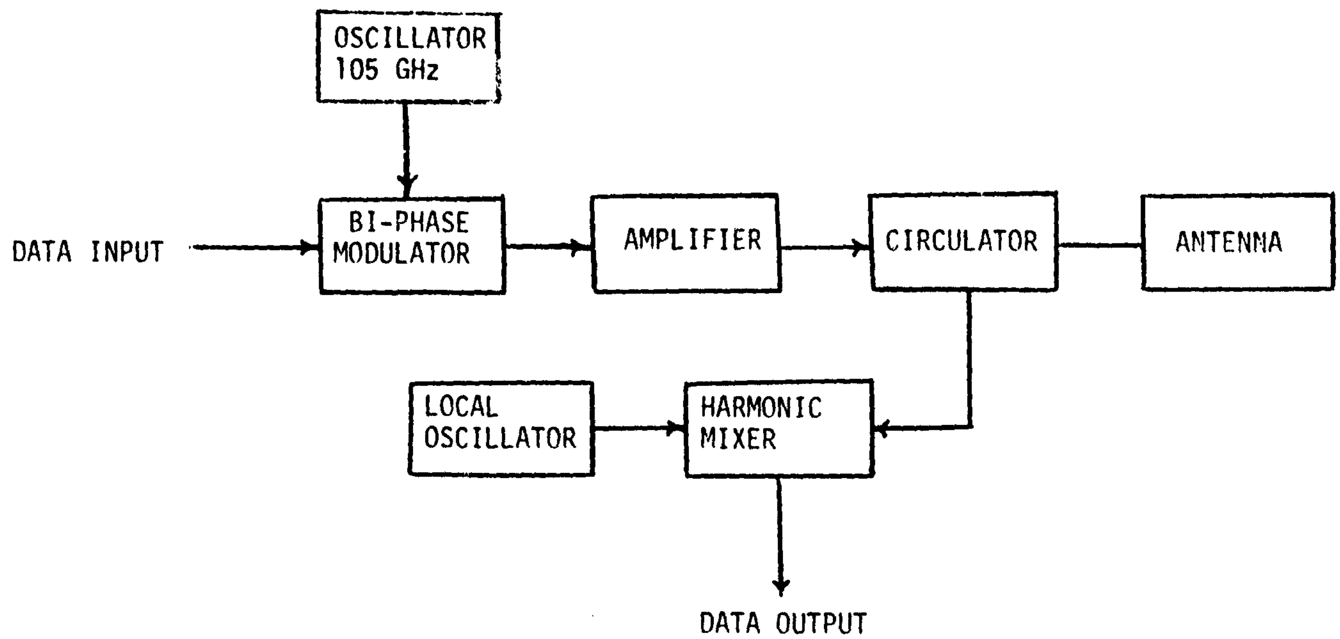



TABLE 1: SYSTEMS PERFORMANCE ANALYSIS WITH 6 INCH ANTENNA

| SYSTEM DESCRIPTION | | | | PATH LENGTH (MI) FOR A BANDWIDTH OF: | | |
|---|---|---|---|---|---|---|
| NO | SOURCE | MODULATOR | MIXER | 1 GHz | 500 MHz | 5 MHz |
| 1 | REFLEX KLYSTRON | PIN BI-PHASE | BAL MIX-IF PREAMP | 50 | 70 | 700 |
| 2 | REFLEX KLYSTRON | PIN BI-PHASE | HARMONIC MIXER | 11 | 15 | 151 |
| 3 | REFLEX KLYSTRON | PIN BI-PHASE | SINGLE ENDED | 34 | 50 | 494 |
| 4 | CW IMPATT SOURCE | PIN BI-PHASE | BAL MIX-IF PREAMP | 18 | 25 | 248 |
| 5 | CW IMPATT SOURCE | PIN BI-PHASE | HARMONIC MIXER | 4 | 5 | 54 |
| 6 | CW IMPATT SOURCE | PIN BI-PHASE | SINGLE ENDED | 12 | 18 | 176 |

An IMPATT amplifier/combiner has been identified which would be suitable for use in the RF subsystem. Kuno and English[2] have developed a technique by which a one or two stage, hybrid-coupled, reflection amplifier could be developed. This device would couple either IMPATT amplifiers or injection locked oscillators together using 3 dB short slot hybrids. For a single stage amplifier operating at 105 GHz we can expect a power output of ~85 mW. For a two stage amplifier in this frequency range, we can expect a power output of ~145 mW. Table 2 contains computed ranges expected for various RF subsystems containing single or dual stage amplifiers. A 200% improvement in range is expected for the single stage amplifier and a 250% improvement in range is expected for the dual stage amplifier. Although an RF amplifier will not be used in the demonstration system for Phase III due to the high cost incurred in developing one, this is a viable source of amplification for future RF subsystem designs.

TABLE 2: SYSTEMS PERFORMANCE ANALYSIS WITH 6 INCH ANTENNA AND C.W. IMPATT SOURCE

| SYSTEM DESCRIPTION | | | | PATH LENGTH (MI) FOR A BANDWIDTH OF: | | |
|---|---|---|---|---|---|---|
| NO. | MODULATOR | AMPLIFIER | MIXER | 1 GHz | 500 MHz | 5 MHz |
| 1 | PIN BI-PHASE | SINGLE STAGE IMPATT | BAL MIX IF-PREAMP | 36 | 49 | 496 |
| 2 | PIN BI-PHASE | SINGLE STAGE IMPATT | HARMONIX MIXER | 8 | 11 | 105 |
| 3 | PIN BI-PHASE | SINGLE STAGE IMPATT | SINGLE ENDED | 24 | 33 | 334 |
| 4 | PIN BI-PHASE | DUAL STAGE IMPATT | BAL MIX-IF PREAMP | 46 | 65 | 648 |
| 5 | PIN BI-PHASE | DUAL STAGE IMPATT | HARMONIC MIXER | 10 | 14 | 139 |
| 6 | PIN BI-PHASE | DUAL STAGE IMPATT | SINGLE ENDED | 31 | 44 | 437 |

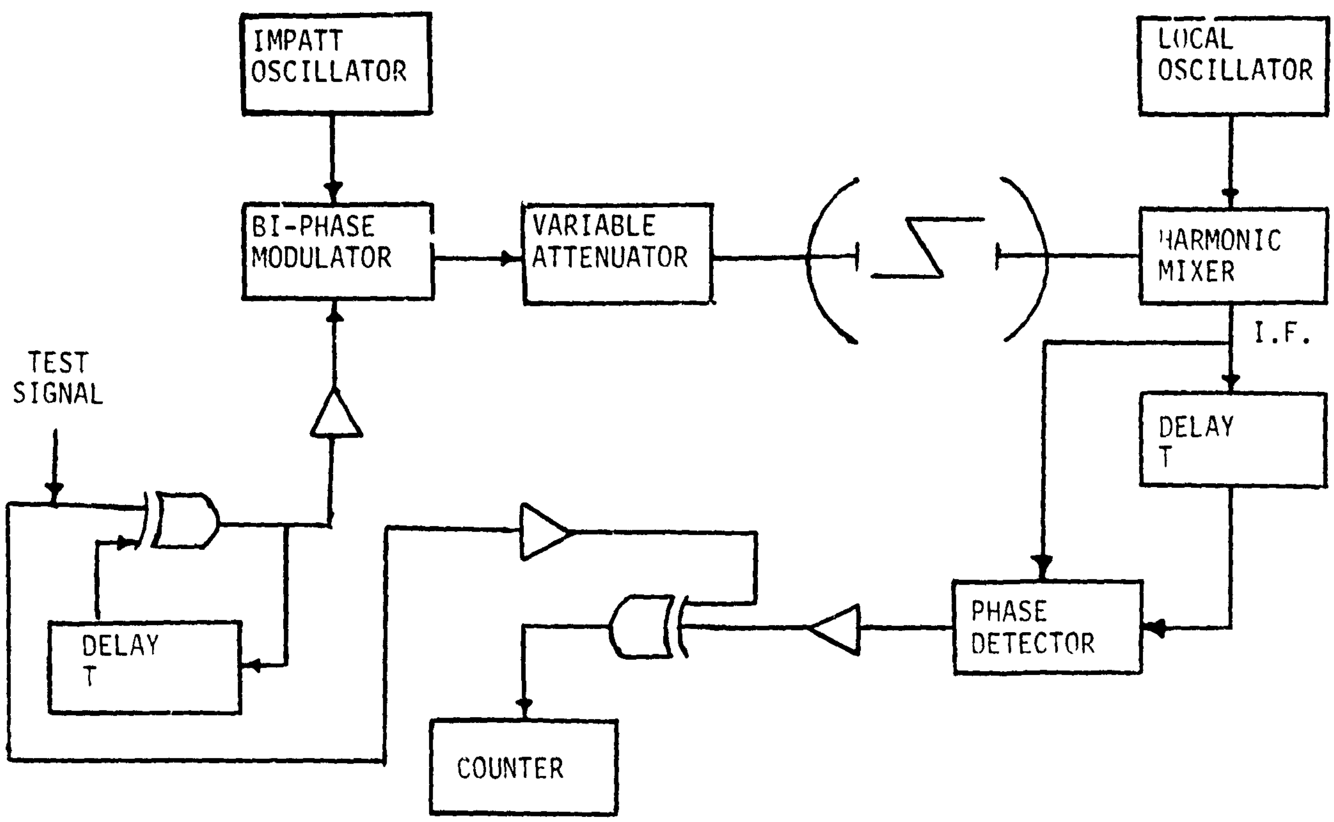
### 2.1.2 Summary of Signal Processing Subsystem Design

The possible techniques for signal processing (generating and monitoring the test signal) have been studied and a design capable of generating the 450 Mb/s test signal required for the Phase III demonstration has been selected.

The test signals will be generated and monitored by Fairchild 100k series ECL (emitter coupled logic) devices. These components offer 500 MHz (typical) clock rates and we have experience with them in our Electronics Subdivision.

A schematic diagram of the planned demonstration system is shown in figure 2. The test signal will be a psuedo-random sequence generated by a Fairchild F100141 shift register operating in a feedback mode with a Fairchild F100107 exclusive-OR gate (shown in figure 3). Our planned modulation format is differential phase shift keying. Therefore a circuit consisting of an exclusive OR-gate and a

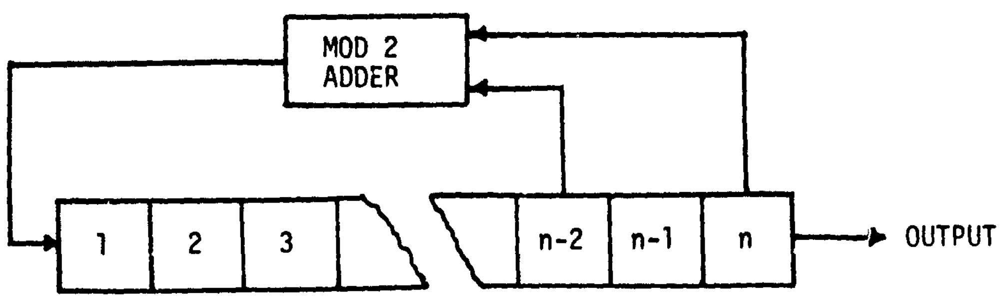
FIGURE 2: SCHEMATIC FOR HIGH DATA RATE DEMONSTRATION



1 bit (approximately 2 ns) delay line will be used to differentially encode the generated test sequence. The ECL logic levels will then be buffered to match the modulation drive requirements.

FIGURE 3: MAXIMAL LENGTH SEQUENCES

| NUMBER OF STAGES | LENGTH OF MAXIMAL SEQUENCE; N | NUMBER OF MAXIMAL SEQUENCES; M | FEEDBACK STAGE CONNECTIONS |
|---|---|---|---|
| 2 | 3 | 1 | 2, 1 |
| 3 | 7 | 2 | 3, 2 |
| 4 | 15 | 2 | 4, 3 |
| 5 | 31 | 6 | 5, 3 |
| 6 | 63 | 6 | 6, 5 |
| 7 | 127 | 18 | 7, 6 |




SHIFT-REGISTER GENERATOR

At the receiver, demodulation of the differential code is accomplished by correlating the encoded data stream with a one bit delayed version of itself. The output of the demodulator will be a bipolar video which must be buffered to the ECL levels. A bit error rate detector will be implemented by comparing the received signal to a hard wire version of the originally generated sequence in an exclusive OR-gate. Bit errors will be counted using an ordinary laboratory counter. Phase matching of the two signals will be accomplished by adjusting the path lengths of the transmitted signal. Error rate versus range will be simulated by attenuating the transmitted signal.

The IF frequency of the receiver is constrained by the harmonic mixer to be in the range of 50 - 500 MHz. Therefore an IF center frequency of 275 MHz will permit the 450 MHz bandwidth required for the data rate goal. The maximum local oscillator frequency allowed by the mixer is 15 GHz. This will be used to minimize the harmonic number, thereby maximizing the efficiency of the mixer. The center frequency of the data signal will therefore be 7 x 15 - .275 = 104.725 GHz.

### 2.1.3 Summary of Alternative Modulation and Demodulation Studies

During Phase I differential bi-phase shift keying was chosen as the modulation scheme for the Ultramicrowave Communications system. However, during Phase II alternate bi-phase modulation-demodulation methods were studied, with the conclusion that several other demodulation concepts could increase the performance of the system. These methods are described below.

#### Bi-Phase Shift Keying Demodulation Studies

A biphase modulated signal can be demodulated by synchronous detection (by mixing it with a reference carrier and low-pass filtering). This is shown figure 4. The difficulty with this approach is that, in a communication system, the reference signal is unavailable at the receiver and must be derived from the received signal itself using one of a number of methods. Two carrier recovery

methods are the squaring loop and the Costas loop. A third method, differential bipolar shift keying (DBPSK) does not recover the carrier itself, but uses the signal phase of the preceding bit period as the reference signal.

FIGURE 4: DEMODULATION OF BIPHASE MODULATED SIGNAL BY SYNCHRONOUS DETECTION

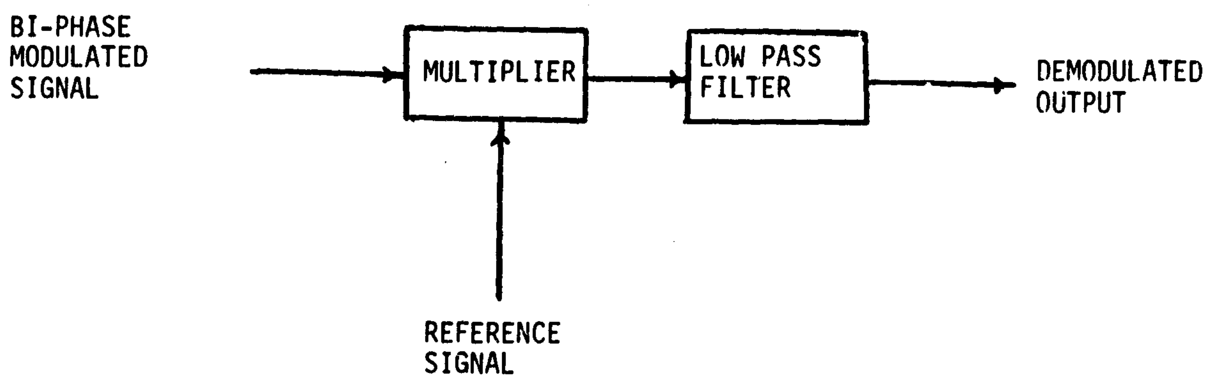
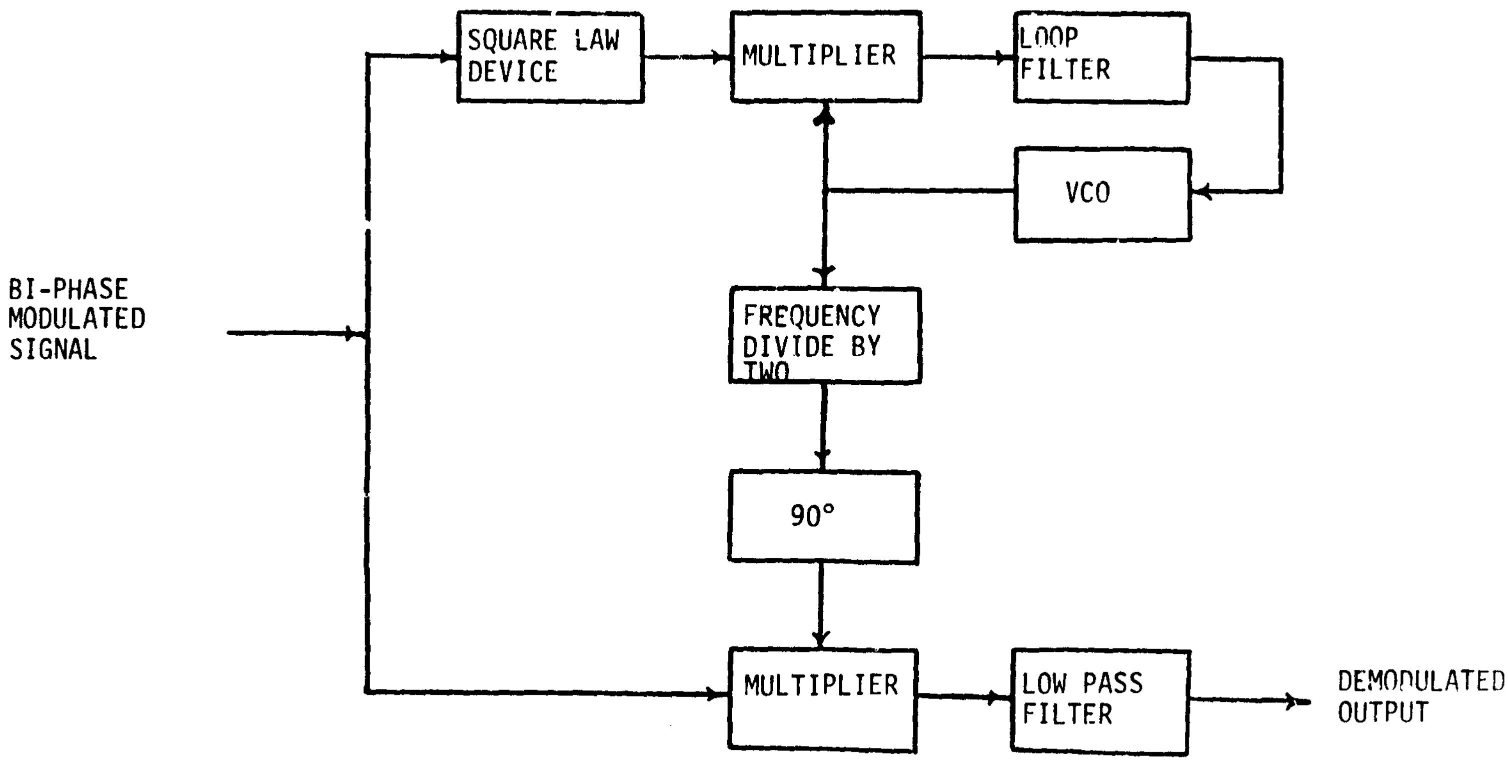



Figure 5 shows a demodulation that uses a squaring loop for carrier recovery[3]. The signal is passed through a square law device which doubles its frequency and removes the phase modulation, then a voltage controlled oscillator is locked to this signal in a phase locked loop. The output of the phase locked loop oscillator is at double the carrier frequency and phase shifted by 90 degrees from the carrier signal, so the signal is passed through a frequency divide by two circuit and a phase shifter before being used as the reference signal in a synchronous detector.

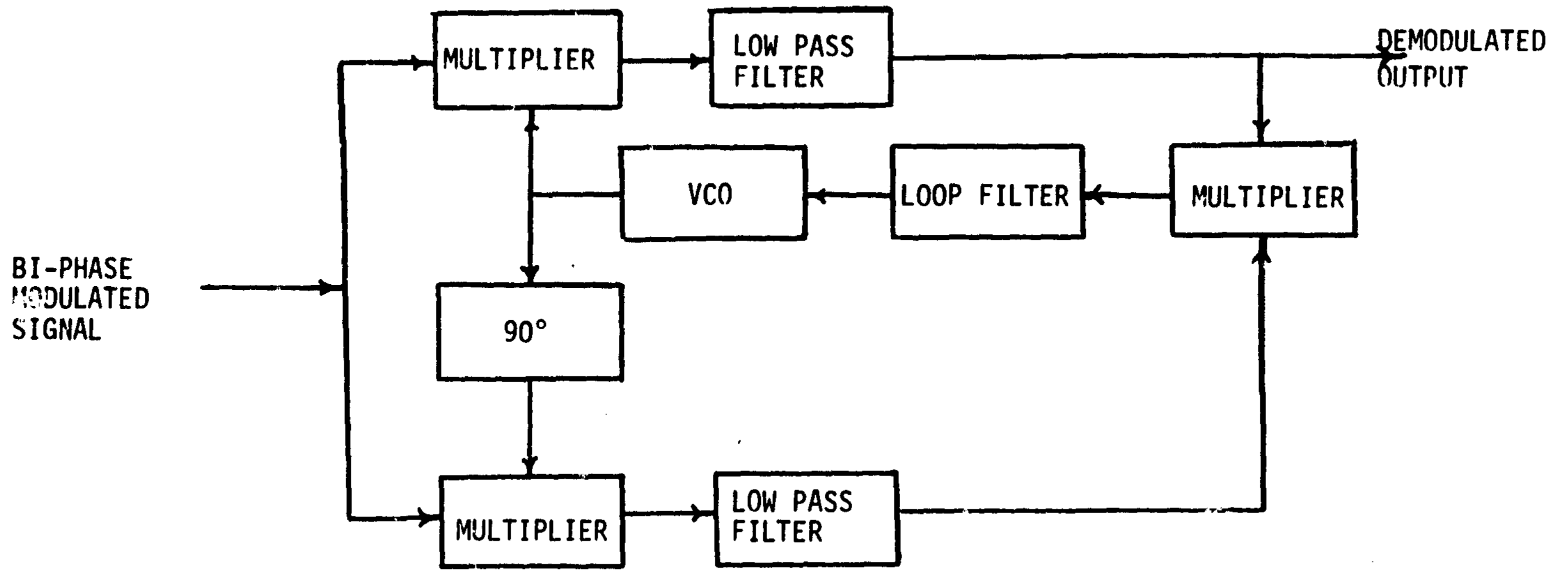
Figure 6 shows a demodulator which uses a Costas loop for carrier recovery[4]. The signal is mixed in a quadrature mixer with the output of a voltage controlled oscillator (VCO). The output of the quadrature mixer is low pass filtered, multiplied, and low pass filtered again, and used to control the frequency of the VCO. The result is a phased locked loop in which the VCO is allowed to lock at either 0 or 180 degrees with respect to the received signal. Since the received

FIGURE 5: DEMODULATOR FOR BI-PHASE MODULATED SIGNAL USING SQUARING LOOP FOR CARRIER RECOVERY




signal bits can have a phase of either 0 degrees or 180 degrees relative to the recovered carrier, the demodulated output is either the desired signal or the negative of the desired signal, an ambiguity which is easily resolved by transmission of a simple reference signal. (The squaring loop also exhibits this ambiguity). Since the loop filter is much narrower than the signal bandwidth, the signal-to-noise performance of the carrier recovery loop is better than that of the signal demodulation itself. Thus clean reference signals can be extracted by the squaring or Costas loops for use in the demodulation process.

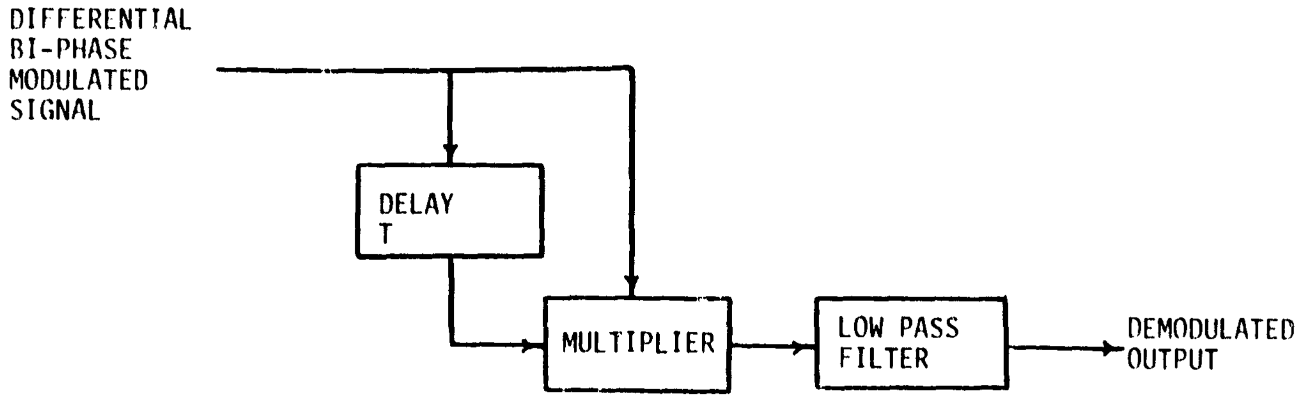
FIGURE 6: DEMODULATOR FOR BI-PHASE MODULATED SIGNALS USING COSTAS LOOP FOR CARRIER RECOVERY




A DBPSK signal does not require recovery of the carrier signal. The signal is specially encoded in the modulation process so that a phase difference of 180 degrees between two successive bit intervals results in the transmission of a digital "1", and no phase difference between two successive bit intervals is the transmission of its complement. Figure 7 shows a block diagram of a DBPSK demodulator [5]. The received signal is applied to a mixer and to a delay line of time delay equal to one bit interval. The output of the delay line is also applied to the mixer. The mixer output is low pass filtered to yield the demodulated output.

For the Ultramicrowave Communications system, direct modulation with a Costas loop demodulator, and a DBPSK system are currently being considered. The DBPSK system is less complex, but constrains the data that is to be transmitted to a fixed bit interval. If data of a different clock rate is to be sent, complex multiplexing and demultiplexing must be performed to transform this data to a rate which can be transmitted by the system. The Costas loop demodulator has no bit

FIGURE 7: DEMODULATOR FOR DIFFERENTIAL BI-PHASE MODULATED SIGNALS




interval constraints so it is more flexible. Any data, regardless of clock rate, can be received by the system (subject to bandwidth requirements, of course). For this reason, the Costas loop seems the best choice for use in a demonstration communication system, where a variety of data types may be transmitted. In addition, the noise performance of a carrier recovery system such as Costas loop is approximately 3 dB better than of the DBPSK system[6]. (In a DBPSK system, bit errors tend to occur in pairs). A Costas loop has one additional advantage in that it is able to demodulate double sided-suppressed carrier (DSB-SC) modulation, of which biphase modulation is a special case. A communication system utilizing DSB-SC modulation could more nearly approach the theoretical channel capacity[7] of the system in terms of bits of information per second.

A double sideband signal can be created at the transmitter with a balanced mixer in place of the biphase modulator used for BPSK. With this approach, modulation could be performed directly with video signals, without the need to convert to digital bits using a high-speed A-D converter and multiplexer. Because of the utility of such a system, further consideration of such an approach is warranted.

Preliminary studies of a Costas loop demodulator indicate that a system with a carrier recovery loop noise bandwidth of 1 MHz is possible that would have a hold-in frequency range of $\pm$10 MHz (that is, the loop could track carrier frequency variations of $\pm$10 MHz, which might occur due to transmitter drift or doppler shift), and a pull-in frequency range of 2.25 MHz (for initial locking). These specifications would probably be adequate for a high data rate communications system.

## 2.2 Summary of Packaging Design

A packaging design of the transceiver mockup was performed by the Electronics Equipment Development Department of MDAC-STL. Size and interconnection requirements of the RF subsystem along with the estimated sizes of non-RF subsystems (signal processing subsystem, power supply subsystem) were used to form the package design.

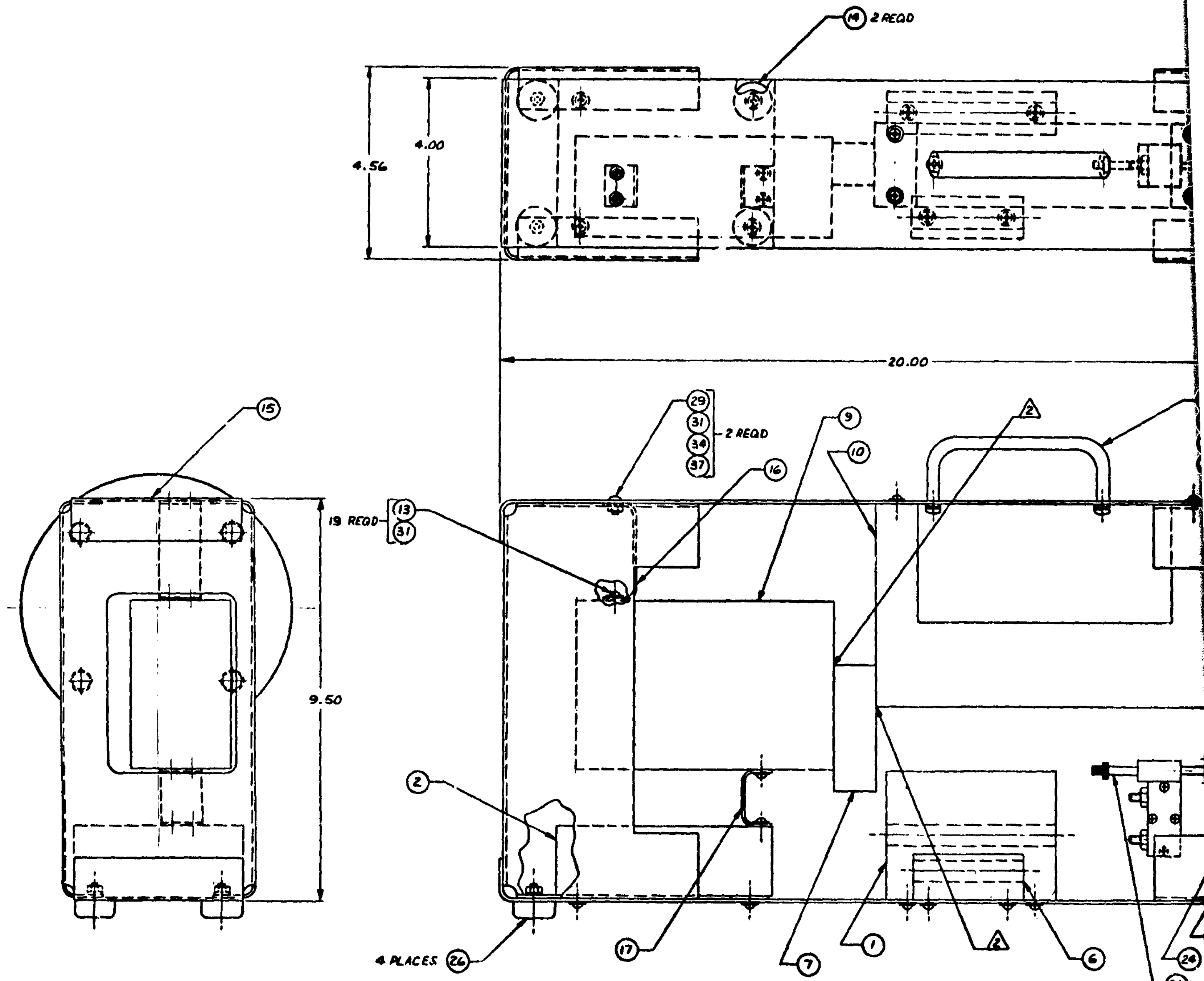
The overall dimensions of the layout are 20.0" x 4.5" x 9.5". This size clearly demonstrates the reduction in volume that can be achieved through the use of ultramicrowave frequencies. The final layout design is somewhat smaller than the preliminary design due to the substitution of the TRG 802-6 antenna for the originally proposed antenna assembly. Figure 8 shows the packaging layouts of the RF subsystem, power supply subsystem, and signal processing subsystem for the transmitter/receiver mockups.

Although the packaging design is essentially complete minor alterations may be necessary in the layout due to changes in the mechanical and electrical specifications of the active hardware. (Most millimeter wave catalog items are just one step removed from laboratory items and therefore are in a state of constant flux).

Figure 8: Receiver/Transmitter Mockup Parts List and Packaging Designs

Receiver Mockup Parts List

| Part Name | Find No. |
|---|---|
| IF Amplifier | 1 |
| Power Supply | 2 |
| Circulator | 3 |
| Local Oscillator | 6 |
| Bi-Phase Mod | 7 |
| CW IMPATT Source | 9 |
| CW IMPATT Amp | 10 |
| Waveguide 90° Bend | 11 |
| Support | 12 |
| Screw, Wood | 13 |
| Washer, Wood | 14 |
| Frame Assy | 15 |
| Brkt-Heat Sink, Upper | 16 |
| Brkt-Heat Sink, Lower | 17 |
| Harmonic Mixer | 21 |
| Parabolic Antenna | 23 |
| Waveguide | 24 |
| Bumper, Rubber | 26 |
| Handle | 27 |
| Bolt, Hex | 28 |
| Screw, Pan Hd | 29 |
| Screw, Pan Hd | 30 |
| Washer, Flat | 31-33 |
| Washer Lock | 34-36 |
| Nut, Hex | 37 |

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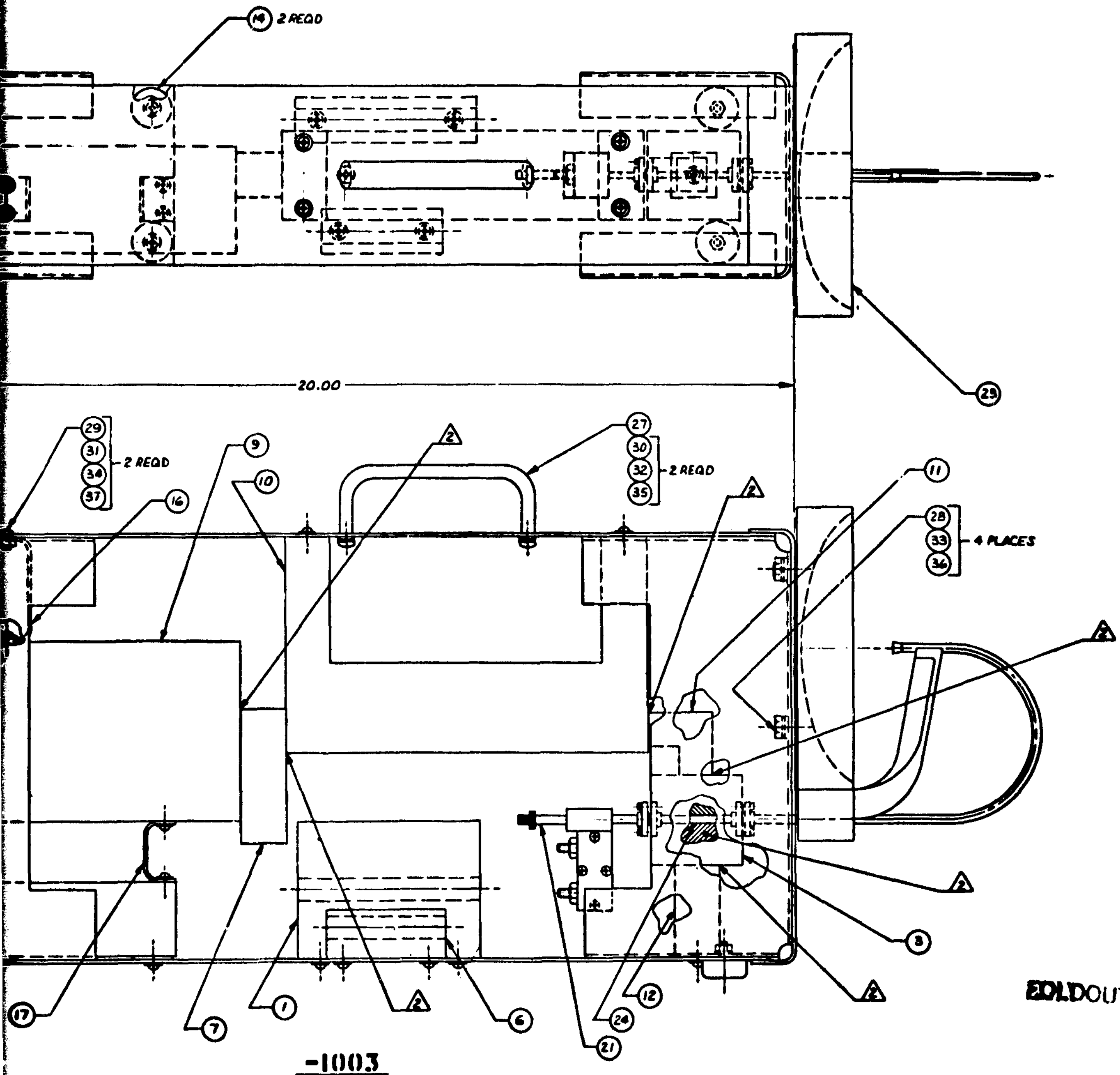
FOLDOUT FRAME




EOLDOUT FRAME 2

ULTRAMICROWAVE COMMUNICATIONS SYSTEM – PHASE II
RECEIVER MOCKUP

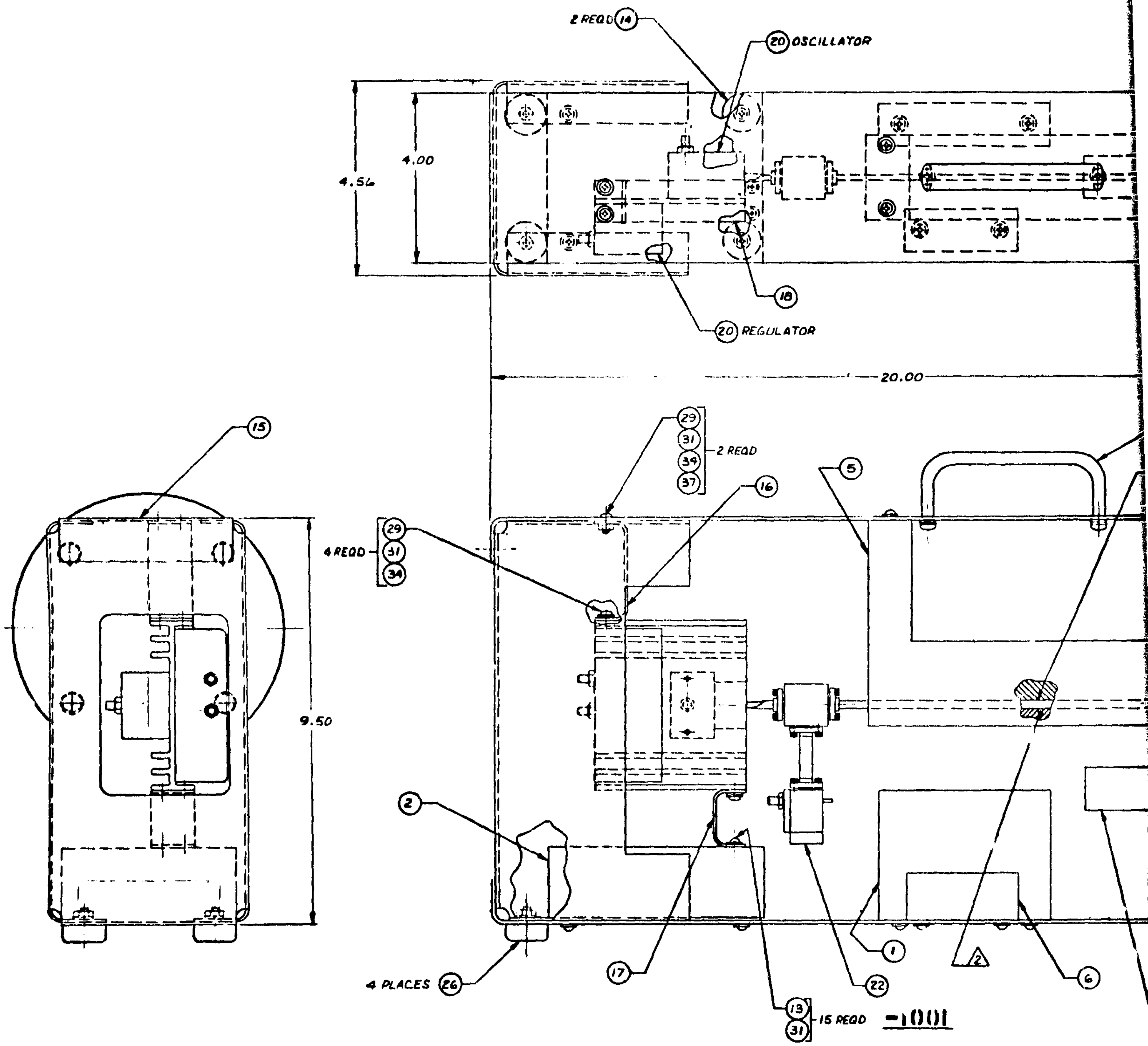
19 SEPT 79 SCALE: 1/1 NO. REQD: 1
SHEET 2 OF 2

 91279RWM REV ___

Transmitter Mockup Parts List

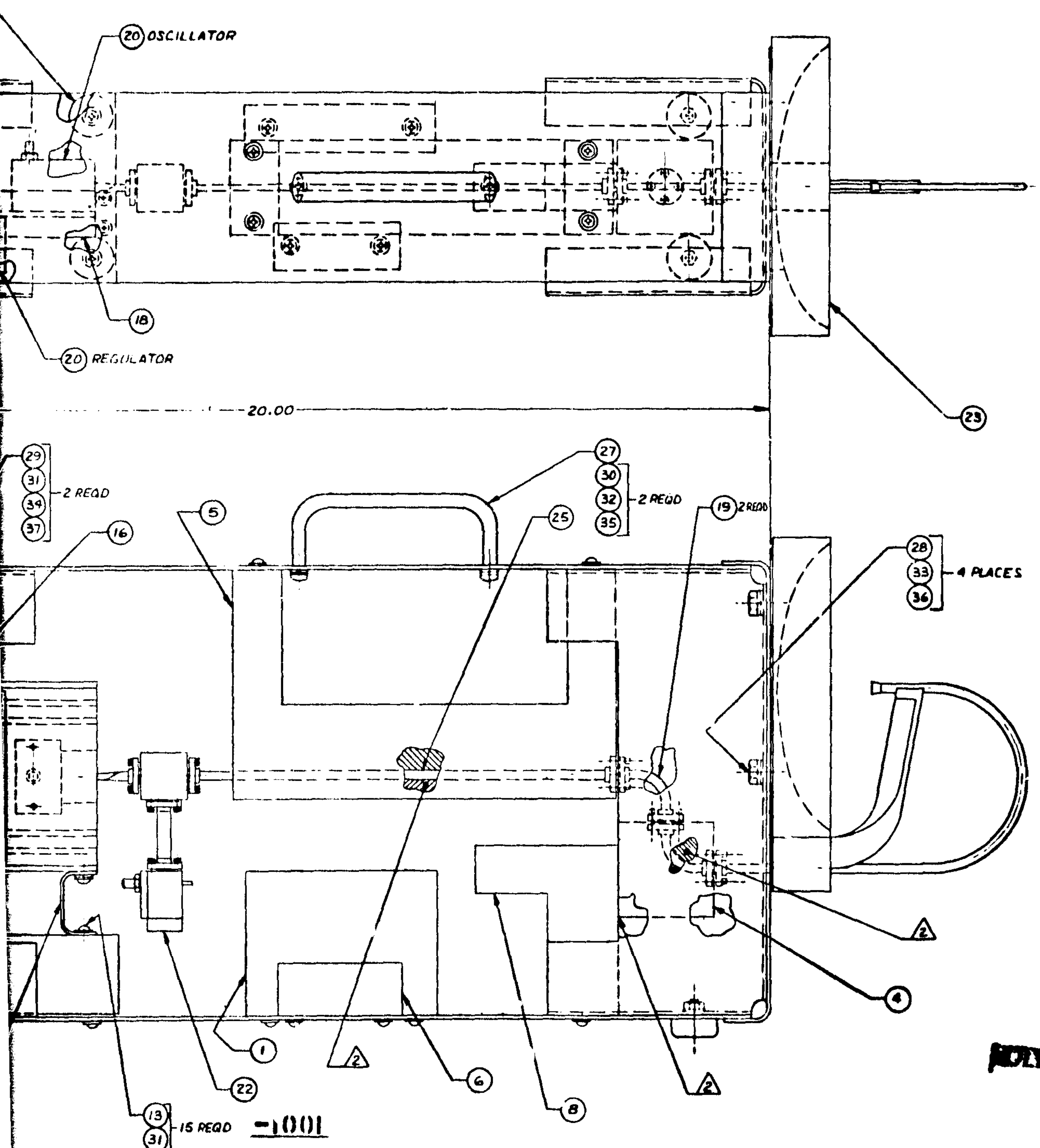
| Part Name | Find No. |
|---|---|
| IF Amplifier | 1 |
| Power Supply | 2 |
| Circulator | 4 |
| CW IMPATT Amp | 5 |
| Local Oscillator | 6 |
| Harmonic Mixer | 8 |
| Screw, Wood | 13 |
| Washer, Wood | 14 |
| Frame Assy | 15 |
| Brkt-Heat Sink, Upper | 16 |
| Brkt-Heat Sink, Lower | 17 |
| Heat Sink | 18 |
| Waveguide, 90° Bend | 19 |
| CW IMPATT Source | 20 |
| Bi-Phase Modulator | 22 |
| Parabolic Antenna | 23 |
| Waveguide | 25 |
| Bumper, Rubber | 26 |
| Handle | 27 |
| Bolt, Hex | 28 |
| Screw, Pan Hd | 29-30 |
| Washer, Flat | 31-33 |
| Washer, Lock | 34-36 |
| Nut, Hex | 37 |

ULTRAMICROWAVE COM




FOLDOUT FRAME 1

MCDONNELL DO




ULTRAMICROWAVE COMMUNICATIONS SYSTEM – PHASE II
TRANSMITTER MOCKUP

12 SEPT 79 SCALE: 1/1 NO. REQD: 1
SHEET 1 OF 2

17 91279RWM REV__

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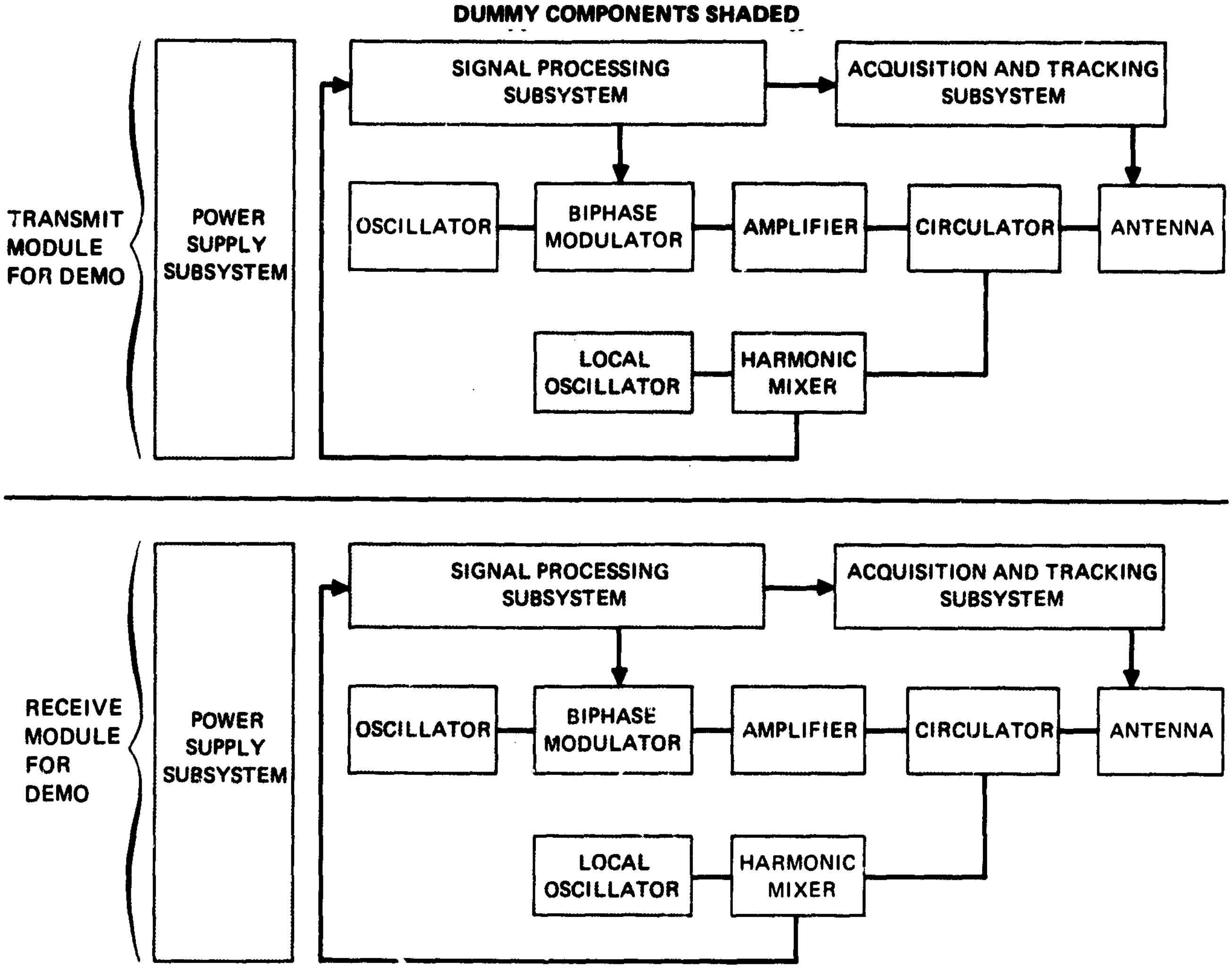
FIGURE 9: FINAL MOCKUP DESIGN




## 2.3 Summary of Mockup Design

A full scale mockup design of a total ultramicrowave communications system was completed and reviewed. Figure 9 shows the RF subsystem designs of the transmitter and receiver. The active RF components for the transmitter mockup are: CW IMPATT source, bi-phase modulator, and parabolic antenna. The active RF components for the receiver mockup are: antenna and harmonic mixer. All dummy components have been designed and drawings suitable for release to the MDAC-STL model shop, for subsequent fabrication, have been produced.

The power supply subsystem will consist of currently available laboratory power supplies. Other components necessary to demonstrate one way communications will be obtained from existing laboratory equipment.

## 2.4 Dual Mode Feasibility

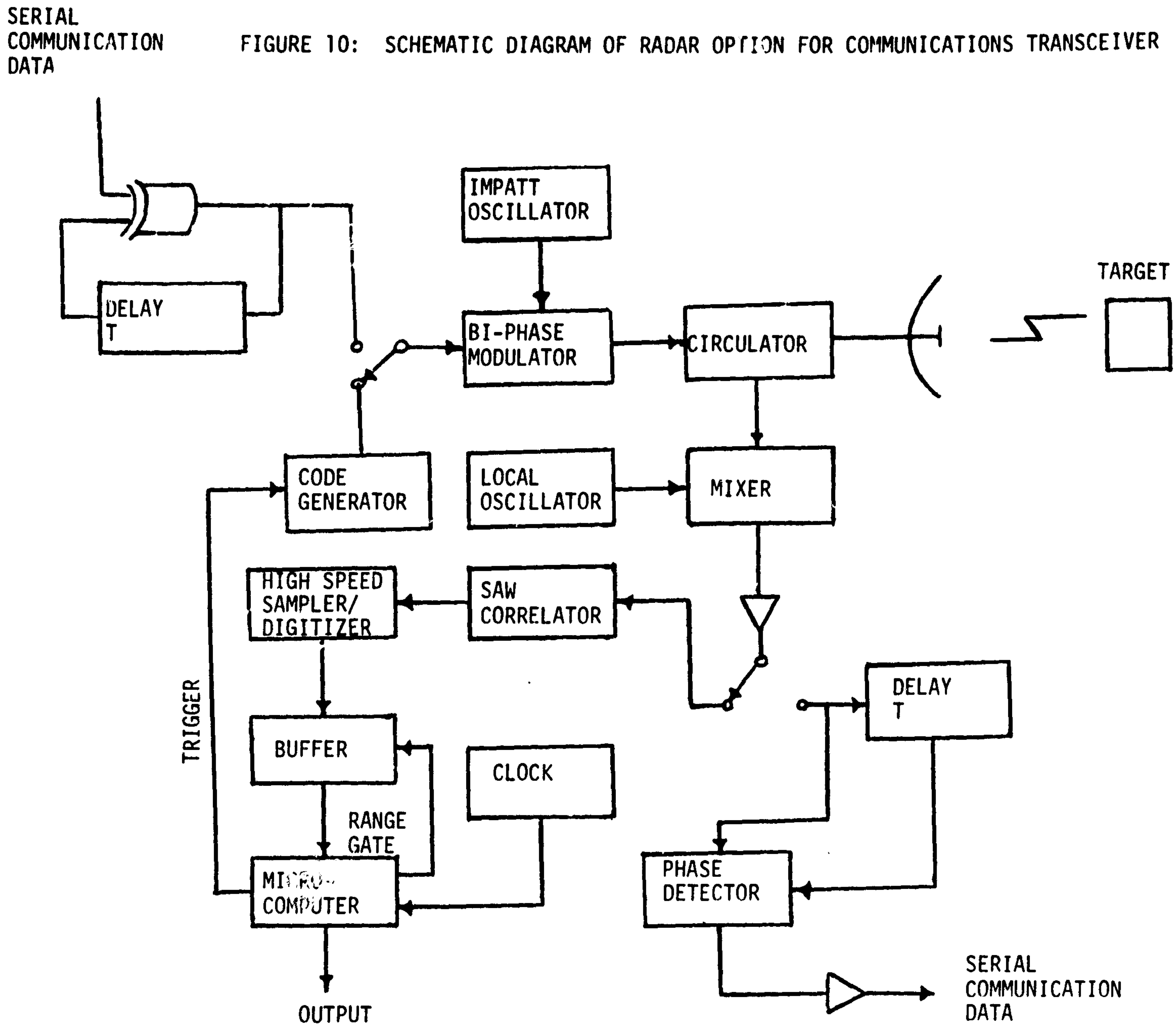
The communications transceiver under development for the ultramicrowave communications system contains all of the RF subsystem components required for a short range radar. The high data rate capability of the communications system implies that high range resolution (on the order of one foot) could be achieved by a radar mode. Also the advantage of having private spectrum allocation for a radar would clearly be attractive. There are many potential applications for the communications system where a high range resolution radar mode would be applicable. e.g., rendezvous and tracking of orbiting payloads, large space structure applications, station keeping, docking, etc.

### 2.4.1 Summary of Preliminary Radar Option Design

In order to minimize the cost of adding a radar option to the communications transceiver no new RF subsystem components need be added. Instead new technologies such as high speed logic, surface acoustic wave (SAW), and microprocessor devices should be utilized. The format to be used for the radar mode is pulse compression of the radar return signal. Pulse compression of the returned signal can be achieved with the use of a bi-phase coded waveform compatible with the bi-phase modulation format of the communications transceiver. This waveform will realize the range resolution capabilities of the wide bandwidth of the ultramicrowave systems. Since all functions required for the radar are implemented at IF frequencies the promise of compactness, low cost and light weight can be realized.

Figure 10 shows a simplified schematic diagram of the radar option. Transmitted signals can be generated by a maximum length sequence generator (a shift register operating in a feedback mode) as shown in figure 3. The received signal

FIGURE 10: SCHEMATIC DIAGRAM OF RADAR OPTION FOR COMMUNICATIONS TRANSCEIVER




will be correlated by a SAW device (the SAW correlator must have the same code as the test signal generator). This output is an analog signal which can be sampled or displayed in time to reveal multiple targets for each pulse. The use of a high speed sampler/digitizer with a digital buffer to match the data requirements of a microprocessor allows tracking of multiple targets on a pulse to pulse basis.

Range rate is available by measuring the doppler frequency shift of the transmitted carrier in a straight forward manner. The use of the ultramicrowave frequencies permits more rapid determination of range rate, since the doppler shift is proportional to transmitted frequency as well as range rate. (For example, the doppler frequency associated with one foot per second is 200 Hz at 100 GHz and only 20 Hz at 10 GHz. To measure these frequencies one would need two or more zero crossings to have a reliable measurement. Clearly, determination of the doppler frequency is 10 times more rapid at 100 GHz than at X-band).

2.4.2 Preliminary Performance Evaluation of Radar Option

The performance of the RF subsystem acting in the radar mode is a function of several parameters. The determination of maximum range is dependent upon transmitter power, antenna gain, target cross section, receiver sensitivity, processing gain, and minimum signal to noise ratio. All of the above parameters were defined in the final report of Phase I (MDC E2059) except for target cross section and processing gain. For a target cross section we will use 10 $m^2$ as the minimum area to be detected. Larger targets (or cooperative targets using a corner reflector) can be detected at greater ranges.

Signal processing gain is primarily dependent upon the type of correlator used. For our initial analysis, we chose a phase coded SAW delay line featuring a code length of 128 chips and a bandwidth of 10 MHz (corresponding to a range resolution of 50 feet). Custom devices are offered with bandwidths to 50 MHz (10 feet). Use of the 128 chip correlator gives a processing gain of 20 dB, which enhances the signal to noise ratio by the same facto (we assumed a 12 dB minimum signal to noise ratio).

The determination of performance involves the use of the above parameters along with the previously determined performance of the communications RF subsystems. The relationship of the various parameters to the radar range is determined as follows:

First we assume our target is in the ar field (Fraunhofer) region of the transmitting antenna. For this to be valid d (distance between target and antenna) must satisfy[8]:

$$d \geq \frac{2D^2}{\lambda}$$

D = antenna diameter (m)
$\lambda$ = carrier wavelength (m)

For our case: $$d \geq \frac{2(.1524\ \mathrm{m})^2}{(2.857 \times 10^{-3}\ \mathrm{m})} \cong 16.26\ \mathrm{m}$$

Therefore for distance $\geq$ 16.26m the radiated power density at a distance, R, from the antenna is:

$$P_d = \frac{P_T G}{4\pi R^2}$$

$P_T$ = power from transmitter
G = antenna gain

For a target of cross section $\sigma$ and distance from antenna, R, the intercepted power is:

$$P_i = \frac{P_T G \sigma}{4\pi R^2}$$

This power is reflected back to the radar system, whose antenna has an effective aperture of:

$$A = \frac{G\lambda^2}{4\pi}$$

Therefore the power received by the antenna after the return trip is:

$$P_a = \frac{P_i A}{4\pi R^2} = \frac{P_T\ G\ \sigma\ G\ \lambda^2}{(4\pi R^2)(4\pi R^2)\ 4\pi} = \frac{P_T\ G^2\ \lambda^2\ \sigma}{(4\pi)^3\ R^4}$$

Using the above argument we can note that, in the communications system, the power received at the receiving antenna is:

$$P_i = \frac{P_T\ G}{4\pi R_1^2}\ \frac{G\lambda^2}{4\pi}$$

where $R_1$ = range of the communications system found in 2.2.1.

Equating this result with the previously obtained result (which implies the minimum detectable signal is the same in both the communications and radar mode) we find:

$$\frac{P_T\ G^2\ \lambda^2}{4\pi R_1^{\ 2}\ 4\pi} = \frac{P_T\ G^2\ \lambda^2\ \sigma}{(4\pi)^3\ R^4}$$

or solving for R we obtain:

$$R = \left(\frac{\sigma}{4\pi}\right)^{1/4} R_1^{\ 1/2}$$

($\sigma$ and R in $m^2$ and m respectively).

In order to include the signal processing gain of the radar receiver we define an $R_1^{\ 1}$ such that:

$$R_1^{\ 1} = R_1\ .\ 10^{S/20}$$

where S = signal processing gain (in dB).

Therefore:

$$R = \left(\frac{\sigma}{4\pi}\right)^{1/4} (R_1\ .\ 10^{S/20})^{1/2}$$

Using this equation Tables 3 and 4 were derived.

TABLE 3: PROJECTED PERFORMANCE (MAXIMUM RANGE) FOR RADAR OPTION AGAINST A $10M^2$ TARGET

| SYSTEM DESCRIPTION | | | | PATH LENGTH (m) FOR A BANDWIDTH OF: | | |
|---|---|---|---|---|---|---|
| NO | SOURCE | MODULATOR | MIXER | 1 GHz | 500 MHz | 5 MHz |
| 1 | REFLEX KLYSTRON | PIN BI-PHASE | BAL MIX-IF PREAMP | 843 | 1000 | 3170 |
| 2 | REFLEX KLYSTRON | PIN BI-PHASE | HARMONIC MIXER | 392 | 466 | 1474 |
| 3 | REFLEX KLYSTRON | PIN BI-PHASE | SINGLE ENDED | 697 | 843 | 2666 |
| 4 | CW IMPATT SOURCE | PIN BI-PHASE | BAL MIX-IF PREAMP | 502 | 596 | 1886 |
| 5 | CW IMPATT SOURCE | PIN BI-PHASE | HARMONIC MIXER | 233 | 277 | 877 |
| 6 | CW IMPATT SOURCE | PIN BI-PHASE | SINGLE ENDED | 415 | 502 | 1588 |

TABLE 4: PROJECTED PERFORMANCE (MAXIMUM RANGE) FOR RADAR OPTION WITH C.W. IMPATT SOURCE, AGAINST A $10M^2$ TARGET

| SYSTEM DESCRIPTION | | | | PATH LENGTH (m) FOR A BANDWIDTH OF: | | |
|---|---|---|---|---|---|---|
| NO | MODULATOR | AMPLIFIER | MIXER | 1 GHz | 500 MHz | 5 MHz |
| 1 | PIN BI-PHASE | SINGLE STAGE IMPATT | BAL MIX-IF PREAMP | 710 | 844 | 2669 |
| 2 | PIN BI-PHASE | SINGLE STAGE IMPATT | HARMONIC MIXER | 329 | 391 | 1237 |
| 3 | PIN BI-PHASE | SINGLE STAGE IMPATT | SINGLE ENDED | 582 | 692 | 2190 |
| 4 | PIN BI-PHASE | DUAL STAGE IMPATT | BAL MIX-IF PREAMP | 811 | 964 | 3049 |
| 5 | PIN BI-PHASE | DUAL STAGE IMPATT | HARMONIC MIXER | 376 | 447 | 1413 |
| 6 | PIN BI-PHASE | DUAL STAGE IMPATT | SINGLE ENDED | 665 | 792 | 2503 |

# 3. PROGRAM PLAN DEVELOPMENT

RF subsystem components identified as being sufficient to demonstrate one-way communications at 105 GHz were ordered from previously identified vendors. A complete assembly plan (including the plans for fabricating dummy components) for the mockup design of the communications system was prepared. Also the test plan (including laboratory equipment requirements) for performance evaluation of the system was completed.

## 3.1 Component Procurement

Purchase orders for the following active RF subsystem components were completed and issued to the respective vendors.

| Description | Part No. | Vendor |
|---|---|---|
| Micrometer Tuned CW IMPATT Source | 47136H-1105 | Hughes Aircraft Company, Electron Dynamics Division |
| Pin Bi-Phase Modulator | 47996H-1000 | Hughes Aircraft Company, Electron Dynamics Division |
| Harmonic Mixer | 47436H-1020 | Hughes Aircraft Company, Electron Dynamics Division |
| (2) 6-inch Prime Focus Antenna | W802-6 | Alpha/TRG Inc. |
| 9-inch Straight Waveguide | W690 | Alpha/TRG Inc. |
| 2-inch Straight Waveguide | W690 | Alpha/TRG Inc. |
| (2) 90 degree H-Plane Bend | W670 | Alpha/TRG Inc. |

## 3.2 System Assembly Plan

The RF subsystem will be constructed by MDAC-STL model shop personnel. The dummy components for each mockup design (transmitter and receiver) will be fabricated from the drawings completed by the Electronic Equipment department. The active hardware will be released to the shop for subsequent integration with the dummy components using connection and constraint requirements previously

developed. The subsystem will then be packaged in a protective case to allow for easy transportation and demonstration.

The signal processing subsystem will be constructed in the Electromagnetics laboratory and then matched with the RF subsystems to produce the communications transmitter/receiver mockups for the Phase III demonstration.

The flexibility of the present assembly and design plans will allow for future realization of a complete communications system by substitution of active components for the dummy components.

3.3 System Test Plan

The test plan of the assembled mockup design and its components has four main objectives: one way communications demonstration, data rate demonstration, size compactness, and component performance analysis.

Products obtained from the previously mentioned manufacturers will be individually tested to determine performance limits and to compare to vendors specifications. This includes ECL as well as millimeter wave components. The RF subsystem will then be assembled, using the plan outlined above, to show size and weight reductions achievable with millimeter wave frequencies. One way communications will be demonstrated using the constructed transmitter/receiver pair. Figure 11 shows an artists' concept of the Ultramicrowave Communications system test. Data rate performance analysis will be conducted and recommendations for the improvement of the maximum obtainable data rate will be made.

Shown below is a partial list of the laboratory equipment needed to perform the tests outlined above:

2 ns delay lines

power supply (50 Vdc/100 mA)

local oscillator

variable attenuator

bi-phase modulator driver

double balanced IF mixers

laboratory counter

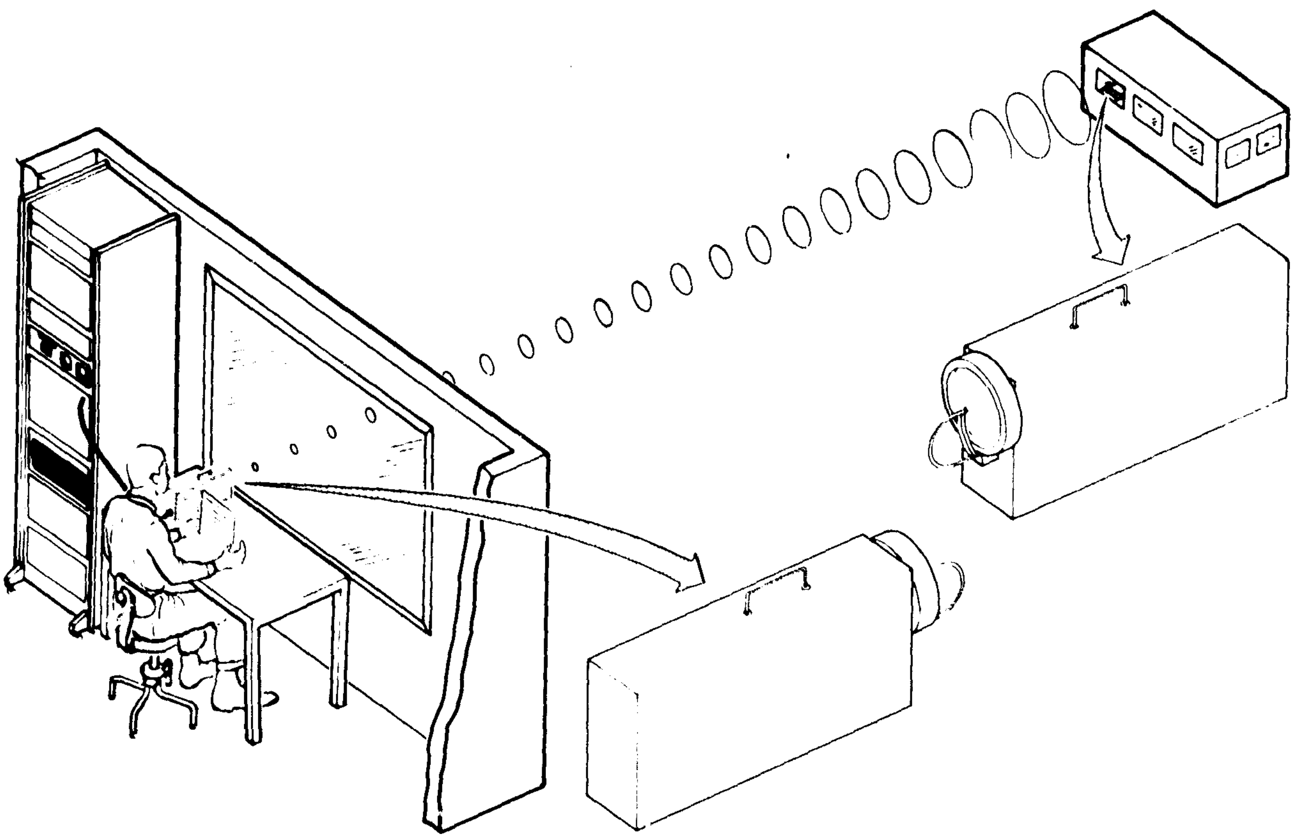
exclusive OR gates

SAW devices

50 - 500 MHz IF amplifiers

It is evident that the performance of the above tests will aid in the adjustment of the range parameters (bandwidth, antenna efficiency, signal to noise ratio, etc.). Clearly, this will permit a more accurate prediction to be made of the performance of the system operating in a space to space mode.

FIGURE 11: **ULTRAMICROWAVE COMMUNICATIONS SYSTEM TEST CONCEPT**



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